

Frequency Down-Converters as Applied to VLBI

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The frequency conversion of a received radio frequency spectrum, from an extra galactic radio source down to video-band frequencies to facilitate data recording or transmission, is quite difficult. A special converter permits this down-conversion while rejecting image-noise "fold over"; however, careful design is required to minimize error contribution to the received information. One possible solution is through the use of combination analog and digital circuitry.

I. Introduction

The frequency down-converters used in very long baseline interferometry (VLBI) systems are necessary to permit recording a segment bandwidth (BW), presently 2 MHz wide or less, from an observed extra-galactic radio source (EGRS) spectrum (Ref. 1). This down-conversion overcomes the incompatibility that exists in obtaining the signal spectrum in a usable form, since a receiving system with adequate sensitivity for detecting extreme weak power levels of the radio sources is at the higher frequency portion of the total spectrum, e.g., S- and X-band, and the magnetic tape recorders, which are available for recording, are in the video band frequencies of approximately 10 kHz to 4 MHz.

The tape recording is a means of obtaining, simultaneously, the radio star signal data at two remotely located receiving stations, which are the terminal ends of a VLBI, and to later bring the two recordings together physically at a processing facility for cross-correlation where the interferometry information is extracted from the radio star observation. An alternate means of getting the data to a common location (to perform the cross-correlation), is to transmit the data via transmission

circuits of NASCOM or directly via a satellite. The down-conversion, at the remote locations, into a narrow band of video frequencies is still a necessity, because the data rate or bandwidth limitations of these transmission circuits places a cost constraint on the operations of a VLBI system.

II. The Problem of Image Spectrum

A fundamental problem in down-converting an RF spectrum to the lower frequency bands is that of image or noise "fold over" into the final desired band. For example, Fig. 1 illustrates a typical super-heterodyne receiver converting a spectrum of 2260 to 2300 MHz to an IF band of 260 to 300 MHz. As shown in the figure, the same IF band can be heterodyned by the same local oscillator from an RF band 1700 to 1740 MHz (called the image band). This is undesirable since any signal from the image band causes a signal-to-noise ratio (SNR) degradation. The particular chosen local oscillator and IF has separated the image-band away from the desired band (the relation: image-band is displaced $2 \times$ IF center frequency, e.g., $2280 - (2 \times 280) = 1720$ MHz) permitting realizable filters to attenuate this image-band.

However, to down-convert a portion of this IF band to video frequencies approaching dc, the above relation of the image-band still exists as shown in the example of Fig. 1b (where a 1-MHz segment BW from 281 to 282 MHz is desired to be downconverted to 3 to 4 MHz) and the resulting “image” band of 274 to 275 MHz cannot be attenuated because of the difficulty of a realizable filter. If the video-band desired is shifted further towards dc, the image band becomes closer to the desired band where finally the image-band is adjacent to the desired band; as shown in Fig. 1c, the image is “folded” into the down-converted dc to video-band range with the second local oscillator situated between these two bands at the IF. This places a burden on realizable filters for rejecting the image band and, hence, other techniques must be utilized.

One technique utilizes the principle of signal cancellation, used for generating single-sideband (SSB) modulation in voice communication application in the reverse order; e.g., a SSB modulator normally takes a baseband signal (in the voice or video range) and modulates a higher RF, and with proper phasing and summation, only the upper sideband is selected (Ref. 2). For the VLBI down-converter application, the “image-band” is cancelled and only the desired band is used. This application has created some semantic problems and expressions such as, “single-sideband demodulator is used to create a base-band signal from the observed radio star spectrum...”; such an expression is confusing to a data communication engineer, since the radio star is considered a wide-band noise-like source.

What is really desired is to translate a segment of the noise-like, RF spectrum (of the radio star) down to video-band frequencies without degradation to the received SNR due to “image-noise fold over.”

This confusion is extended further by another VLBI application, termed Δ VLBI, which observes alternately the EGRS signal and a spacecraft transmitted radio frequency signal for navigation purposes (Ref. 3). The spacecraft spectrum is either a carrier only (line spectra) signal for narrow-band Δ VLBI, or a sinewave modulated carrier signal for wideband Δ VLBI, where the sidebands (also line spectra) are widely spaced about the carrier frequency (typically ± 20 MHz).

During the spacecraft signal observation, the requirement is to measure the phase information contained in the extreme sideband frequencies to determine the group delay of the transmitted signal. The method used is to down-convert each sideband and, through low pass filtering, obtain essentially a narrowband width signal, near dc, without image noise fold over.

The procedure is not demodulation to extract the baseband signal (since the baseband signal would be 20 MHz), but the translation of a particular segment of the spacecraft RF spectrum similar to that achieved with the EGRS spectrum.

III. Sideband Spectrum Down Conversion

Figure 2 is a schematic of an analog circuit that achieves the desired down-conversion and functions as follows:

The top multiplier multiplies the input signal spectrum $V_{sig} = V_S \sin(\omega_S \rightarrow \omega_{LS})t + \sin(\omega_S \rightarrow \omega_{US})t$ with the local oscillator $V_{LO} \sin \omega_{LO}t$. Where $\sin(\omega_S \rightarrow \omega_{LS})t$ is designated a spectrum band of frequencies from the spectrum center frequency, ω_S , to a lower sideband range ω_{LS} (here sideband is meant the spectrum band on the lower side of the spectrum center frequency), and the second term of the signal as the upper sideband range of frequencies.

The top multiplier output is:

$$\begin{aligned} V_{TO} &= V_S [\sin(\omega_S \rightarrow \omega_{US})t + \sin(\omega_S \rightarrow \omega_{LS})t] \\ &\quad [V_{LO} \sin \omega_{LO}t] \\ &= \frac{V_{LO} V_S}{2} \left\{ \cos[(\omega_S \rightarrow \omega_{US}) - (\omega_{LO})]t - \cos[(\omega_S \rightarrow \omega_{US}) + (\omega_{LO})]t \right. \\ &\quad \left. + \cos[(\omega_S \rightarrow \omega_{LS}) - (\omega_{LO})]t - \cos[(\omega_S \rightarrow \omega_{LS}) + (\omega_{LO})]t \right\} \quad (1) \end{aligned}$$

Let $\omega_{LO} = \omega_S$; then $(\omega_S \rightarrow \omega_{US}) > \omega_{LO}$ and $(\omega_S \rightarrow \omega_{LS}) < \omega_{LO}$. Therefore,

$$\begin{aligned} V_{TO} &= \frac{V_{LO} V_S}{2} \left\{ \cos(0 \rightarrow \omega_{US})t - \cos(2\omega_S \rightarrow \omega_{LS})t \right. \\ &\quad \left. + \cos(0 \rightarrow \omega_{LS})t - \cos(2\omega_S \rightarrow \omega_{LS})t \right\} \quad (2) \end{aligned}$$

It is simple enough to filter out the higher frequency spectrum while preserving the down-converted spectrum, thus:

$$V_{TO} = \frac{k V_S}{2} [\cos(0 \rightarrow \omega_{US})t + \cos(0 \rightarrow \omega_{LS})t] \quad (3)$$

The amplitude term is $k V_S/2$ where $V_{LO} > V_S$ by design so that the incremental conductance of the mixer is controlled

as a definite function of time by the local oscillator voltage and has a parameter that varies as a fixed function of time relative to the signal; in other words, the resultant by-product amplitude is proportional to the signal amplitude with a fixed conversion loss accountable by k .

The bottom multiplier output becomes,

$$V_{BO} = V_S [\sin(\omega_S \rightarrow \omega_{US}) t + \sin(\omega_S \rightarrow \omega_{LS}) t] \\ [V_{LO} \cos(\omega_{LO}) t] \\ = \frac{V_{LO} V_S}{2} \left\{ \sin[(\omega_S \rightarrow \omega_{US}) - (\omega_{LO})] t + \sin[(\omega_S \rightarrow \omega_{US}) + (\omega_{LO})] t + \sin[(\omega_S \rightarrow \omega_{LS}) - (\omega_{LO})] t + \sin[(\omega_S \rightarrow \omega_{LS}) + (\omega_{LO})] t \right\}$$

or

$$V_{BO} = \frac{kV_S}{2} \left\{ \sin(0 \rightarrow \omega_{US}) t + \sin[-(0 \rightarrow \omega_{LS})] t \right\} \\ = \frac{kV_S}{2} [\sin(0 \rightarrow \omega_{US}) t - \sin(0 \rightarrow \omega_{LS}) t] \quad (4)$$

The resultant outputs would represent two sideband spectra "folded-over" each other and converted down to dc, where the edge of the sideband spectrum components are equal or $\omega_{US} = \omega_{LS}$, if they were to be combined now.

But, the top multiplier output signal is instead shifted -90 deg before either being summed or subtracted with the bottom multiplier output signal; hence, Eq. (3) becomes

$$V'_{TO} = \frac{kV_S}{2} [\sin(0 \rightarrow \omega_{US}) t + \sin(0 \rightarrow \omega_{LS}) t] \quad (5)$$

and Eq. (4) becomes

$$V'_{BO} = \frac{kV_S}{2} [\sin(0 \rightarrow \omega_{US}) t - \sin(0 \rightarrow \omega_{LS}) t] \quad (6)$$

after passing through the two respective phase shift and delay networks.

Adding the two outputs, $V_{SUM} = kV_S \sin(0 \rightarrow \omega_{US}) t$ is the upper sideband spectrum only. This is followed by a low-pass filter (LPF) whose "break" frequency is equal to ω_{US} and actually sets the resultant spectrum band, or channel bandwidth, that is down-converted down to video band.

In similar fashion the subtracted resultant is

$$V_{SUBTRACTED} = kV_S \sin(0 \rightarrow \omega_{LS}) t$$

and is the lower sideband without the "image fold-over" from the upper sideband spectrum.

An "all-pass" network to achieve a constant -90-deg phase shift with constant amplitude of the down-converted frequencies over the video band is a nonrealizable network, so iterative network sections have been used to synthesize the requirement (Ref. 4).

The resultant amplitude and phase-versus-frequency characteristics of a typical complete down-converter circuit are shown in Fig. 2 and it is seen that the phase is nonlinear versus frequency. When the phase is magnified, and differenced, from an average curve, it results in ripples across the band as shown in Fig. 2d.

The resulting phase ripple of a segment BW (of the total span bandwidth) as it is down-converted through the various channels (minimum of three) to determine the group delay of the radio star signal, is an error contributor to this determination. A wider span bandwidth tends to reduce this effect, and the nature of the signal's continuous type spectrum produces an average phase delay value through each individual channel.

For example, Fig. 3 illustrates how segment bandwidths are obtained from the RF span bandwidth (with an intermediate channel used, not shown, to resolve the ambiguity of the 2π radians/cycle phase information between the extreme end channels) and used in determining the group delay. This technique is called bandwidth synthesis (Ref. 5). The low-end channel results in a net phase delay of -1680 deg comprised of the total accumulated phase shift from the microwave, RF, IF, and down-converter circuits and similarly, the upper end channel results in a phase shift of -320 deg.

The group delay through the receiving subsystem is determined by the difference in the end channel phase shifts divided by the frequency separation of the channels, or

$$\tau_{group} = \frac{\phi_2 - \phi_1}{f_2 - f_1} \times \frac{1}{360} \frac{\text{deg}}{\text{cycle}} \frac{\text{deg}}{\text{sec cycle}}, \text{ seconds}$$

For the example shown the group delay is 94 ns. Now if each channel contains only 1-deg peak-to-peak phase ripple (most present down-converter phase shift networks are in the range of 5 to 10 deg peak-peak) and if the upper channel frequency phase terminates on the upper +1 deg and lower end on the lower -1 deg, then the second derivative of the slope results in

$$\sigma_{\tau_g} = \frac{d^2\phi}{dt^2} = \frac{1}{360 \times 40 \times 10^6}$$

and produces 69.4 ps or ≈ 2 cm equivalent error as a first-order approximation. This is illustrated in Fig. 4.

The continuous spectrum nature of the EGRS signal tends to average out this error and it becomes apparent how a wider span BW and the nature of the signal obviate this error source; however, as the precision of VLBI observation of radio stars is desired or when a line spectra spacecraft signal with doppler effect is observed for Δ VLBI, then it can be seen that the ripple becomes a large error contributor.

This is compounded by the fact that the various channels both at a given station and the complementary station would be hard to match or even to predict or control under environmental and operational conditions.

Various schemes to reduce the phase ripple can be applied, such as using resistance-capacitance bridge circuits and optimum pole-zero pair locations; however, the need for component precision values, matching, and resulting instability are undesirable.

A method that appears feasible in reducing the phase-ripple is to utilize digital techniques to accomplish the second -90-deg phase delay of the preceding circuit. For example, accomplish the first frequency multiplication of the input signal with quadrature components of the local oscillator, as shown in Fig. 2; however, in place of the delay and -90-deg phase shift delay network, following the mixing substitute low-pass filters with a "break-frequency" matched to the sampling frequency, which is used in two following analog-to-digital converters in place of the sum and subtraction circuit.

The LPF is used to reduce "aliasing" by complementing the sampling rate to satisfy the Nyquist criterion, and is chosen to be a higher value than the final LPF segment bandwidth channel value and thus introduce negligible phase error contribution. Also, it filters the higher frequency by products.

The digital signals can be processed to produce the desired -90-deg phase delay digitally where it is more feasible to

optimize the pole-zero location and ensure repeatability and stability.

Similarly, after the unwanted sideband image "foldover" is eliminated, a following LPF bandwidth using digital technique determines the final channel BW.

It is expected that the phase ripple resulting from this analog-digital combination can achieve values approaching 1/2 deg peak-to-peak.

IV. Down Conversion of a Monochromatic Source

Consider a typical observation of a monochromatic source such as a spacecraft signal as performed during Δ VLBI, and ignore noise (receiver T_{op} , and galactic background); the received signals at the two stations can be represented as

$$V_1(t) = A_1 \cos 2\pi f_{S_1} t \quad (7)$$

and

$$V_2(t) = A_2 \cos 2\pi f_{S_2} (t - \tau_{geo}) \quad (8)$$

where $A_1 A_2$ is the amplitude of the received signals f_{S_1} and f_{S_2} at antennas 1 and 2, respectively; τ_{geo} is the geometric time delay of the signal wave front to antenna 2 after passing antenna 1, and is related to the orientation of the baseline and source signal direction.

The signal is first heterodyned to an intermediate frequency by the first local oscillator, as shown in Fig. 1:

$$\begin{aligned} V_{if_1}(t) &= [A_1 \cos 2\pi f_{S_1} t] [L_1 \cos (2\pi f_{LO_1} t - \phi_1)] \\ &= \frac{A_1 L_1}{2} \cos [2\pi (f_{S_1} - f_{LO_1}) t + \phi_1] \end{aligned} \quad (9)$$

where

f_{LO_1} is the first local oscillator frequency at station 1.

L_1 is the amplitude of the first local oscillator at station 1 and is made larger than the signal amplitude, $L_1 > A_1$.

ϕ_1 represents the phase shift due to the receiver local oscillator and mixer-amplifier.

The band pass characteristics of the IF amplifier rejects the sum term and amplifies the difference term, while the a RF band-pass filter rejected the image frequency, f_{image} (i.e., $\cos 2\pi f_{image} t \times \cos 2\pi f_{1LO_1} t = f_{if}$) as shown in Fig. 1.

With the local oscillator amplitude being greater than the signal amplitude, the first IF signal becomes

$$V_{if_1}(t) = k_1 A_1 \cos [2\pi f_{if_1} t + \phi_1] \quad (10)$$

where k_1 accounts for the mixer conversion loss and $2\pi f_{if_1} t = 2\pi(f_{S_1} - f_{1LO_1})t$.

For the second station, the first IF signal becomes

$$\begin{aligned} V_{if_2}(t) &= [A_2 \cos 2\pi f_{S_2} (t - \tau_{geo})] \\ &\times [L_2 \cos (2\pi f_{1LO_2} t - \phi_2)] \\ &= \frac{A_2 L_2}{2} \cos 2\pi [f_{S_2} (t - \tau_{geo}) - (f_{1LO_2} t - \phi_2)] \\ &\quad + \frac{A_2 L_2}{2} \cos 2\pi [f_{S_2} (t - \tau_{geo}) + (f_{1LO_2} t - \phi_2)] \\ &= k_2 A_2 \left\{ \cos 2\pi [(f_{S_2} - f_{1LO_2}) t - f_{S_2} \tau_{geo} + \phi_2] \right. \\ &\quad \left. + \cos 2\pi [(f_{S_2} + f_{1LO_2}) t - f_{S_2} \tau_{geo} - \phi_2] \right\} \quad (11) \end{aligned}$$

Following the IF amplifier and bandpass filter it becomes,

$$V_{if_2}(t) = k A_2 \cos 2\pi [f_{if_2} t - f_{S_2} \tau_{geo} + \phi_2] \quad (12)$$

where $2\pi f_{S_2} \tau_{geo}$ is the geometric time delay associated with the received radio frequency as shown by Eq. (8) and is typically a slowly varying function of time due to the earth's rotation, or instantaneously is a constant phase in conjunction with the IF signal in Eq. (12).

The final down-conversion to video band is achieved by setting the second local oscillator (see Fig. 5) slightly below the intermediate frequency to avoid shifting the signal to dc

and to accommodate doppler shifts. Adopting the notation of Section III, Eq. (1), the first station IF signal, e.g., (10) multiplied by the second LO

$$\begin{aligned} V_O(t) &= k A_1 \left\{ \cos [2\pi (f_{if_1} \rightarrow f_{US}) t + \phi_1] \right. \\ &\quad \left. + [\cos 2\pi (f_{if_1} \rightarrow f_{LS}) t + \phi_1] \right\} \\ &\times [V_{2LO_1} \sin 2\pi f_{2LO_1} t] \quad (13) \end{aligned}$$

where the second local oscillator frequency is less than the IF, $f_{2LO_1} < f_{if_1}$, and the upper sideband edge is greater than the IF, $f_{if_1} < f_{US}$ with the sideband range containing the IF signal, i.e., $(f_{2LO_1} \rightarrow f_{US})$ brackets $\cos (2\pi f_{if_1} t + \phi_1)$. See Fig. 5a.

Continuing through the down-conversion process, the summed output then rejects the lower sideband image spectrum and passes the upper spectrum with the resultant video band as

$$V_{v_1}(t) = [\sin 2\pi(o \rightarrow f_{US}) t] \quad (14)$$

which contains the RF signal as,

$$V_{v_1}(t) = A_{v_1} \sin (2\pi f_{v_1} t + \phi'_1) \quad (15)$$

where ϕ'_1 is the phase delay modified by the down converter circuits (see Fig. 5b).

Similarly, for the down-converted signal from the second station,

$$V_{v_2}(t) = A_{v_2} \sin 2\pi [f_{v_2} t - f_{S_2} \tau_{geo} + \phi'_2] \quad (16)$$

For the EGRS spectrum the same process results except the final down-converted signal is a continuous spectrum from $o \rightarrow f_{US}$, or equivalently a segment of the radio frequency from $f'_S \rightarrow f_{US}$ where $f'_S = f_{1LO} + f_{2LO}$ or the sum of the first and second local oscillator frequencies.

These signals are later cross-correlated together (Ref. 5) and processed further with the delay τ_{geo} , and delay rate $\dot{\tau}_{geo}$, the basic VLBI information data.

References

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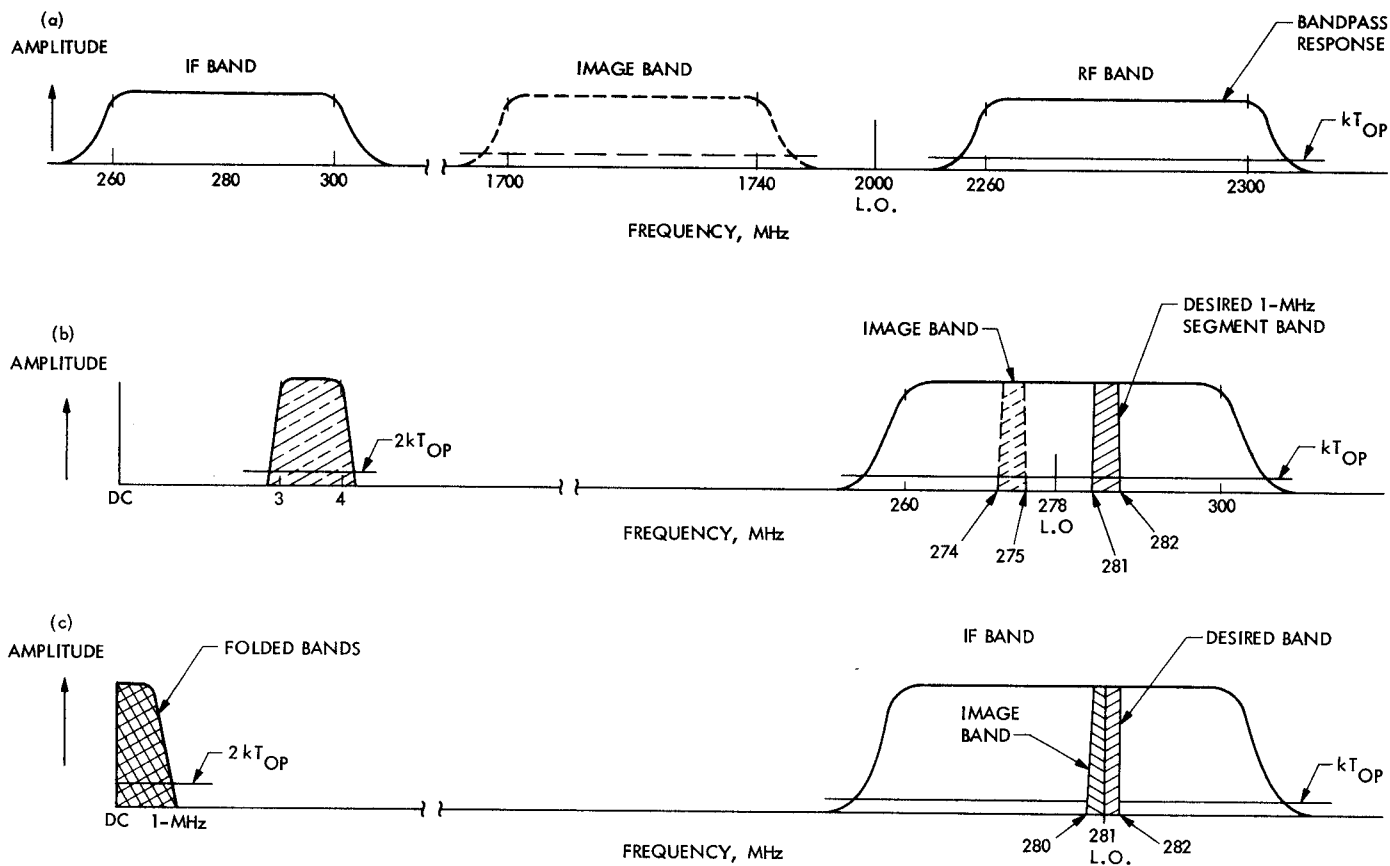


Fig. 1. Image band frequencies resulting from frequency heterodyning, down-conversion: (a) first conversion with rejected image band; (b) second conversion with nonrejected image because of nonrealizable filter; (c) limiting condition where band is to dc when L.O. frequency equals RF band center

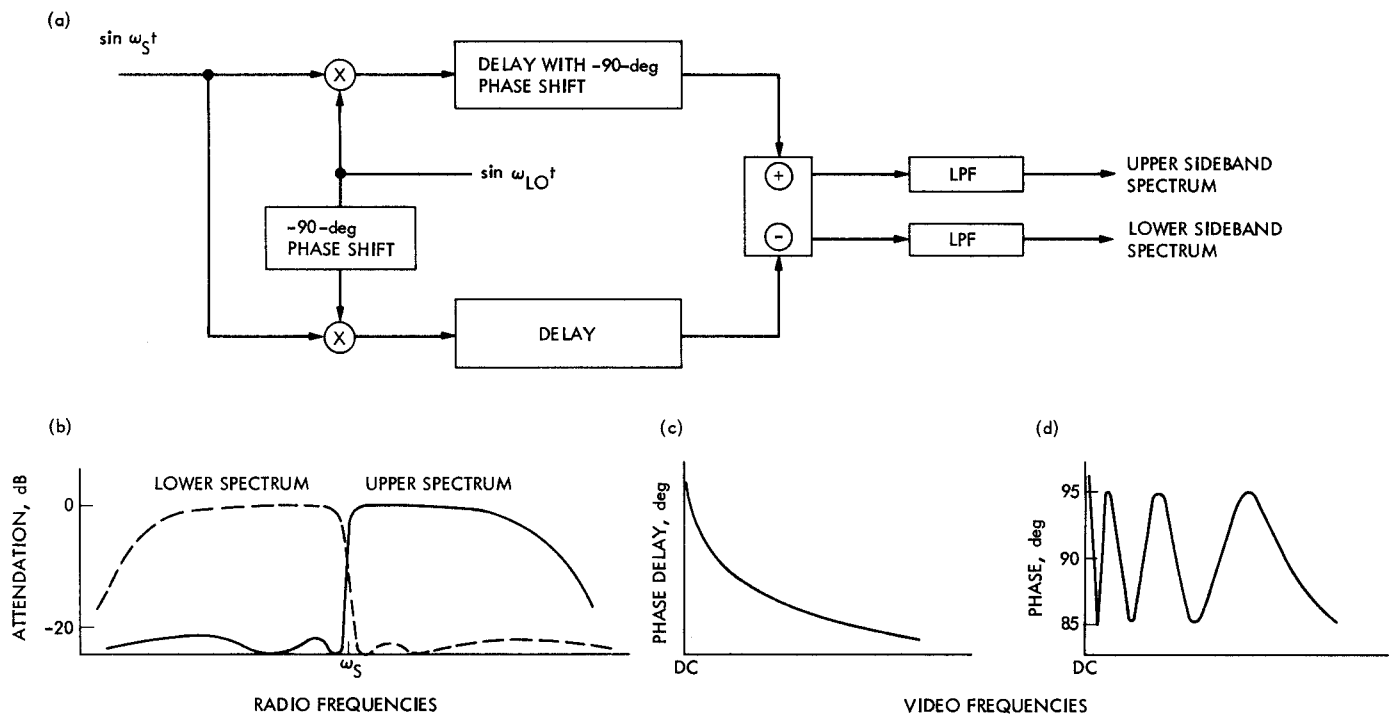


Fig. 2. Frequency down-converter and characteristics: (a) analog circuit; (b) equivalent amplitude response at RF; (c) phase response at video frequency; (d) phase ripple response

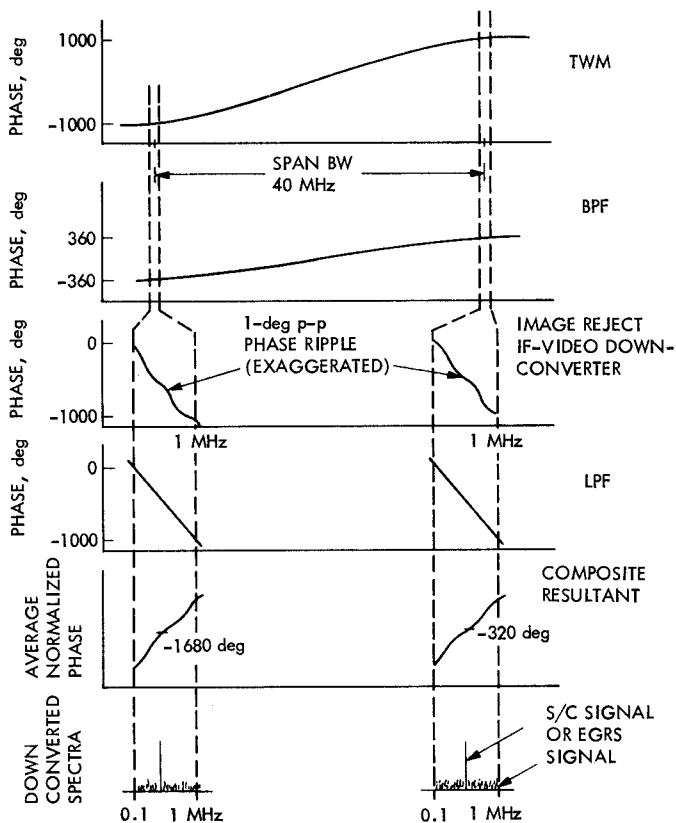
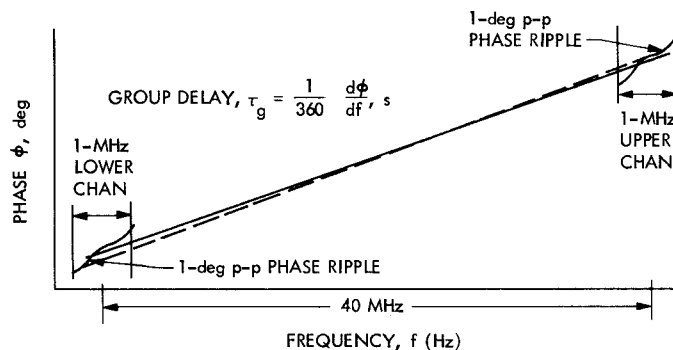


Fig. 3. Resultant down-converted channels



ESTIMATED ERROR IN TIME DELAY MEASUREMENT:

$$\sigma_{\tau_g} = \frac{d^2\phi}{df^2} = \frac{1}{360 \times 40 \times 10^6} = 69.4 \text{ ps} \approx 2 \text{ cm}$$

= GROUP DELAY ERROR PER STATION (USING ASSUMED VALUE EXAMPLE)

Fig. 4. Simplified group delay error evaluation

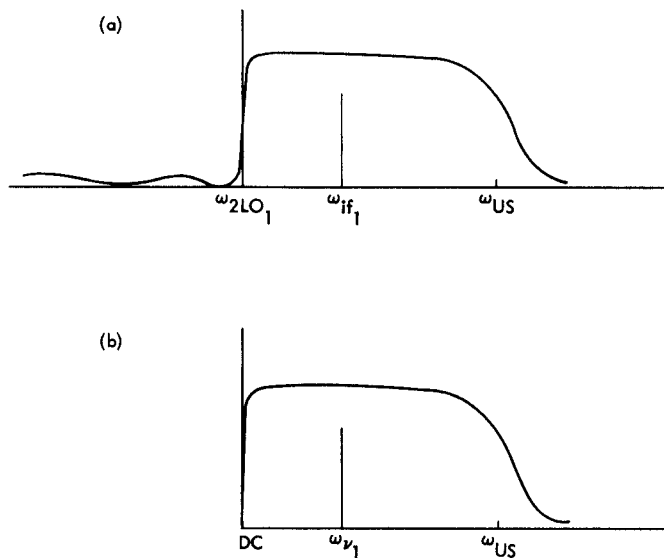


Fig. 5. Down-conversion of a spacecraft line spectra signal: (a) Equivalent amplitude bandpass response at IF where second L.O. is set to less than signal frequency; (b) down-converted signal to video band with resultant RF signal greater than dc and low-pass filter response